28

TRANSMISSION CHARACTERISTICS OF PULSE-SLOPE-MODULATED SIGNALS THROUGH BAND-LIMITED SYSTEMS

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ABSTRACT. Distortion, crosstalk and noise characteristics of P.S.M. have been determined. Maximum harmonic distortion is about 2 per cent for slow cut-off rate of the medium and about 5% for sharp-cut-off filters. Slicers introduce more distortion and non-linearity in modulation. Crosstalk ratios are better than those in P.A.M. for slow cutt-off rate and improvement in crosstalk may be affected by simultaneous introduction of h.f. and l.f. cut-offs. Output S/N ratios show considerable threshold effect and are approximately proportional to the square root of the video bandwidth.

I. INTRODUCTION

The important parameters of a pulse are its amplitude, duration, phase, frequency and slope of the leading and trailing edges. In P.A.M., P.L.M., P.P.M. and P.F.M. systems, the slope is preferably maintained constant at a very high value, whereas in Pulse-slope-modulation (Das, 1954), the slope is varied in accordance with the modulating signal, keeping other parameters constant. The modulating signal is recovered by differentiation and 'Box-car' demodulation

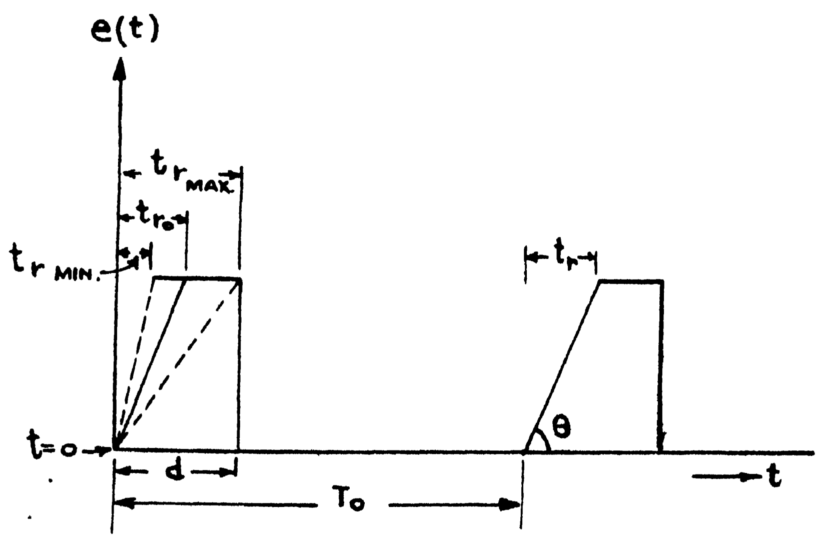

Fig. 1(a). P. S. M. signal. Ch. 1.

5

of the slope-modulated pulses. The frequency spectrum (Das, 1955) of the P.S.M. signal is given by Fig. 1(a) & 1(b):

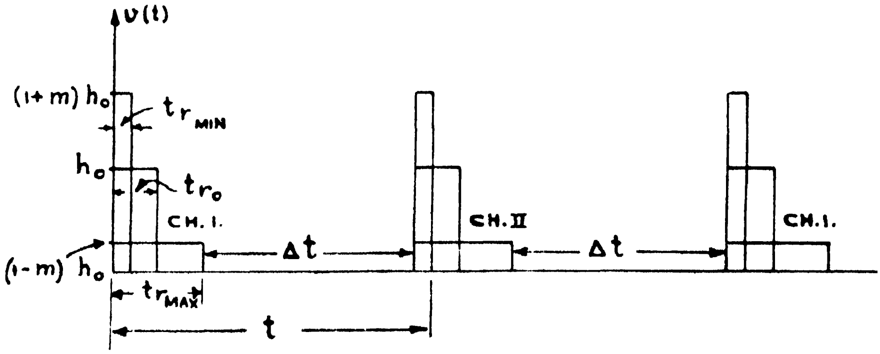

Fig. 1(b) Differentiated signal. Ch. I and Ch. II.

$$F(t) = \frac{Ad}{T_0} - \frac{At_{r_0}}{2T_0(1+m\sin\phi)} + \frac{2A}{T_0}\sum_{n=1}^{\infty}\frac{\sin \omega n(d-t)}{\omega n}$$

$$+\frac{2A}{T_0}\sum_{n=1}^{\infty}\frac{(1+m\sin\phi)}{t_{r_0}\omega^2 n^2}\left[\cos\omega n\left\{t-\frac{t_{r_0}}{(1+m\sin\phi)}\right\}-\cos\omega nt\right] \quad \ldots \quad (1)$$

where,

t_{r_0} = mean risetime

m = modulation index $= \dfrac{t_{r_{max}} - t_{r_{min}}}{t_{r_{max}} + t_{r_{min}}}$

$\omega = \dfrac{2\pi}{T_0}$

$E\sin\phi$ = modulating voltage.

The differentiated signal consists of a pulse train containing both amplitude modulation and inverse width modulation, as is seen from

$$F'(t) = \frac{2A}{T_0}\sum_{n=1}^{\infty} -\cos\omega n(t-d)$$

$$+\frac{2A}{T_0}\sum_{n=1}^{\infty}\frac{(1+m\sin\phi)}{t_{r_0}\omega n}\left[\sin\omega n\left\{\frac{t_{r_0}}{(1+m\sin\phi)}-t\right\}+\sin\omega nt\right]\ldots \quad (2)$$

The inverse width modulation is however cancelled by the use of the 'Box-car' demodulator and the audio output is due only to the amplitude modulation of the differentiated pulses.

Usefulness of a transmission system is determined by the effects of non-ideal circuits and limited bandwidth on its various characteristics—specially audio distortion, crosstalk and noise. These have been discussed here for a P.S.M. system. Since the slope distortion due to limited bandwidth occurs before the 'Box-car' circuit, the effect of both amplitude modulation and inverse width modulation has to be considered for the purpose of determining harmonic distortion and crosstalk. We can generally assume that the system is linear and passive up to the differentiator and on this basis, calculate the theoretical harmonic distortion and crosstalk in the system. For determining the effect of noise, the slope variation only due to noise pulses is calculated for different bandwidths.

The experimental set-up used for the determination of various characteristics had the following specifications (Fig. 2) :

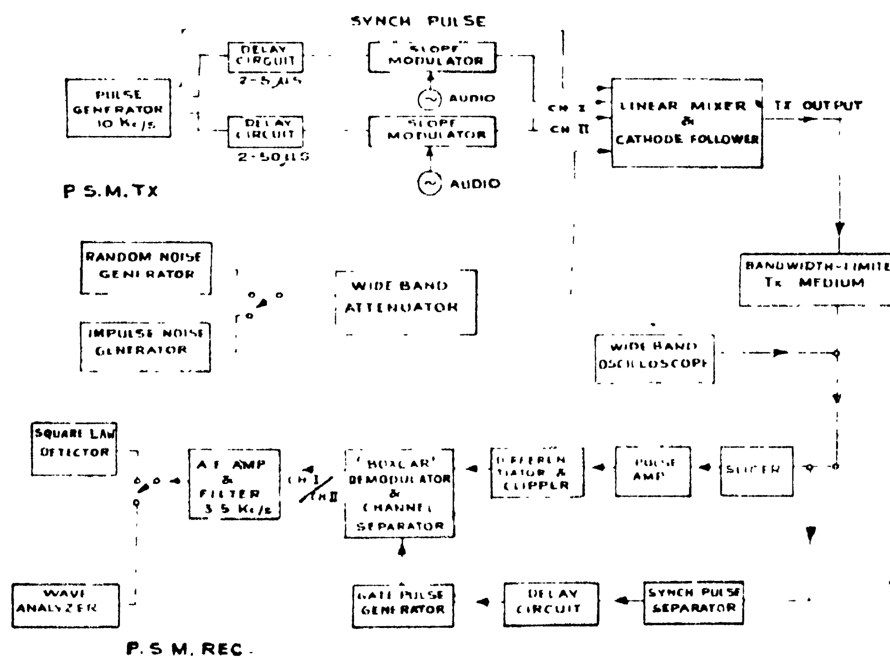

Fig. 2. Experimental set-up—P.S.M. Transmitter and Receiver.

P.R.F. $= 10$Kc/s. $= 1/T_0$

Pulse duration $d = 2-15\ \mu$ sec.

Mean risetime $t_{r_0} = 1-5\mu$ sec.

Delay between Ch. I and Ch. II $= 2-50\ \mu$ sec.

System bandwidth $= 1$ Mc/s.

Audio frequency passband = 0 to 3.5 Kc/s

A.F. modulating frequency = 1 Kc/s.

The modulation was linear for an audio volume range of 40 db. The total harmonic distortion in the received audio output was 2.0% with the peak detector and 0.8% with the 'Box-car' demodulator for a modulation range of 35 db. The inherent crosstalk and noise in the system was approximately 65 db below the signal level.

II. DISTORTION

Harmonic distortion occurs due to non-ideal differentiators, high frequency cut-off, low frequency cut-off and due to the use of slicers for elimination of noise. An ideal differentiator gives an output amplitude equal to τ_1/t_r, where τ_1 is a constant. But with a simple R C differentiator, the amplitude of the differentiated pulses is given by

$$V(t) = \frac{\tau_1}{t_r}\left[1-\exp\left(-t/\tau_1\right)\right] - \frac{\tau_1}{t_r}\left[1-\exp\left\{-\frac{(t-t_r)}{\tau_1}\right\}\right] \cdot U(t-t_r) \quad \ldots \quad (3)$$

where $U(t-t_r)$ = Unit step function starting at $t = t_r$, and τ_1 = RC of the differentiator.

It is thus seen that the output signal does not reach its peak value immediately and with large τ_1, the peak value will not be reached at all for small values of t_r. Due to the associated inverse width modulation, the sharper pulses, corresponding to the peaks of the modulating voltage, will be more attenuated than the wider pulses and the resulting audio output will have its peaks flattened. The use of a peak detector then will give rise to pronounced second harmonic distortion. However, with a 'Box-car' pulse-lengthener circuit, the narrow gate pulse is arranged to occur at $t \leqslant t_{r_{min}}$ and the amplitude of the output pulses becomes proportional to $(1/t_r)$. With gate pulses narrower than the differentiated pulses, the harmonic distortion is very small even for large values of τ_1, as shown in Fig. 3. The experimental total distortion for the peak detector, also shown in Fig. 3, agrees sufficiently with the theoretical values calculated from Eq. (3).

A transmission medium with 6db/octave high frequency cut-off rate, simulated by a simple RC-lowpass filter, gives a differentiated output:

$$V(t) = \frac{\tau_1}{t_r}\left[1 - \exp\left(-\frac{t}{\tau_2}\right)\right] - \frac{\tau_1}{t_r}\left[1-\exp\left\{-\frac{(t-t_r)}{\tau_2}\right\}\right] \cdot U(t-t_r \qquad (4)$$

where

τ_1 = differentiation constant; τ_2 = RC of the lowpass filter.

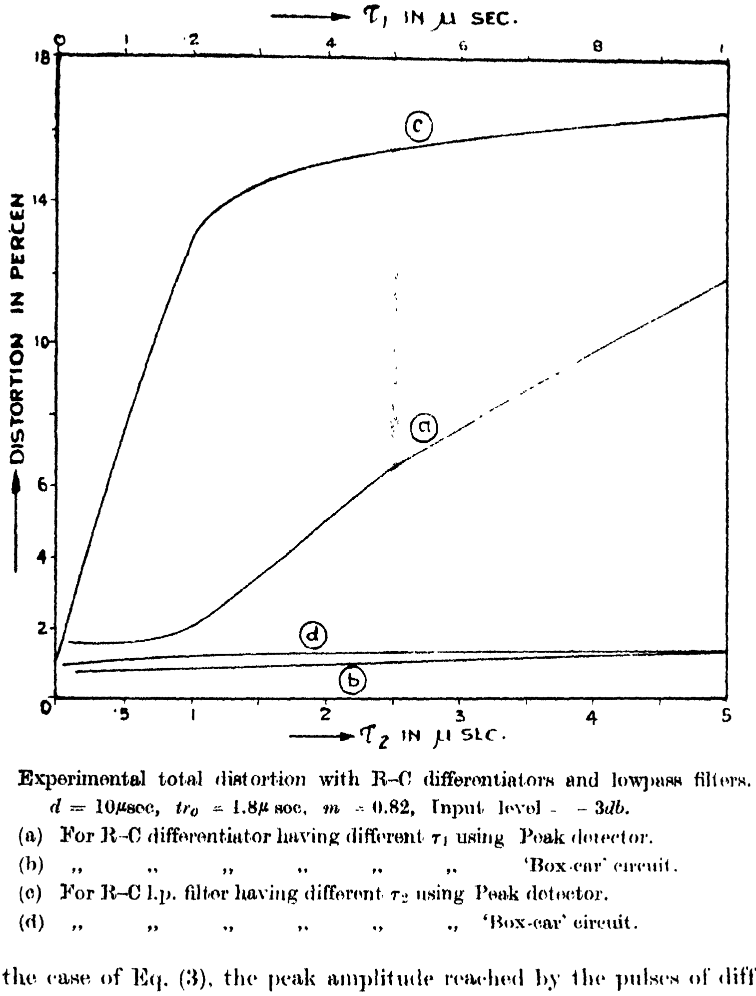

Fig. 3. **Experimental total distortion with R–C differentiators and lowpass filters.** $d = 10\mu$sec, $tr_0 = 1.8\mu$ sec, $m = 0.82$, Input level − −3db.

(a) For R–C differentiator having different τ_1 using Peak detector.

(b) ,, ,, ,, ,, ,, ,, 'Box-car' circuit.

(c) For R–C l.p. filter having different τ_2 using Peak detector.

(d) ,, ,, ,, ,, ,, ,, 'Box-car' circuit.

As in the case of Eq. (3), the peak amplitude reached by the pulses of different widths will not be proportional to τ_1/t_r and with a peak detector, there will be considerable second harmonic distortion. But with a 'Box-car' circuit, arranged to gate at $t \leqslant t_{rmin}$, the distortion is very much minimised, as is seen in Fig. 3.

With transmission media, having sharp high frequency cut-off, the pulse response is oscillatory and considerable distortion occurs with a peak detector. With the 'Box-car' demodulator circuit, the gate pulse is arranged to occur before the oscillation starts; even then, the amplitude of the output pulses is not strictly proportional to τ_1/t_r and there is some distortion in the audio output. The experimental results, as shown in Fig. 4, have been obtained with a variable cut-off electronic filter having 18 db/oct. and 36 db/oct. slopes.

The effect of low frequency cut-off, having sharp as well as slow rate of cut-off, is rather small. Theoretically, the peak amplitude of the differentiated pulses occurs at $t = 0$, and the audio output is distortionless. But with higher cut-off frequency, there is a trailing-edge overshoot (Bhattacharyya, 1953) with distorted

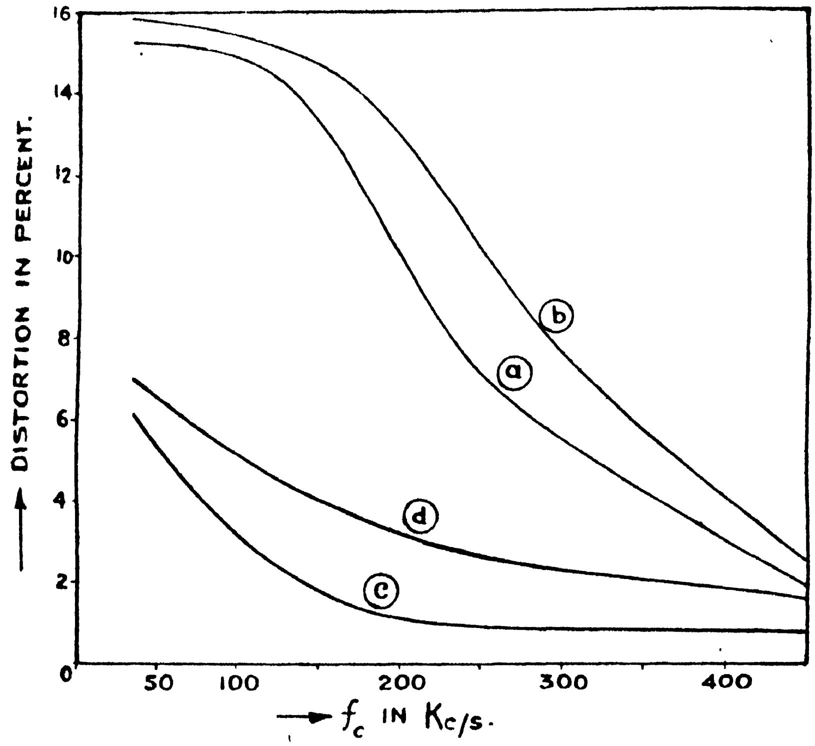

Fig. 4. Experimental total distortion with lowpass electronic filter.
$d = 10\mu$ sec, $tr_0 = 1.8\mu$sec, $m = 0.82$; Input level $= -3db$.
(a) With $18db$/oct cut-off using peak detector.
(b) With $36db$/oct ,, ,, ,, ,,
(c) With $18db$/oct ,, ,, 'Box-car' circuit.
(d) With $36db$/oct ,, ,, ,, ,,

modulation on it. With a peak-detector, this gives rise to certain distortion in the audio output, whereas, the 'Box-car' circuit nullifies the effect of the trailing-edge modulation and the output has very little distortion—approximately 1% only for all practical bandwidths.

A slicing circuit introduces some more non-linearity in modulation and harmonic distortion in the audio output. From noise considerations, the slicing level is generally maintained at half the height of the received pulses. Due to slicing, the differentiated pulses are time-displaced and even after 'Box-car' demodulation, some amount of distorted pulse-length-modulation occurs. However, on further analysis, it is found that the overall distortion due to this P.L.M. is always less than 2%.

The distortion caused by the lowpass band-limited system in the sliced output is rather serious. For pulses with small t_r, the half level is reached much after t_r and the corresponding slope is less than that attained by the unsliced distorted pulses. This makes it necessary to lower the slicing level such that the slicing time $t_s \leqslant t_{r_{min}}$. Fig. 5 shows the nature of the distortion obtained at optimum slicing levels corresponding to the different values of τ_2 and f_c. For highpass filters, the slicer circuits do not contribute to any further distortion, as has been verified experimentally.

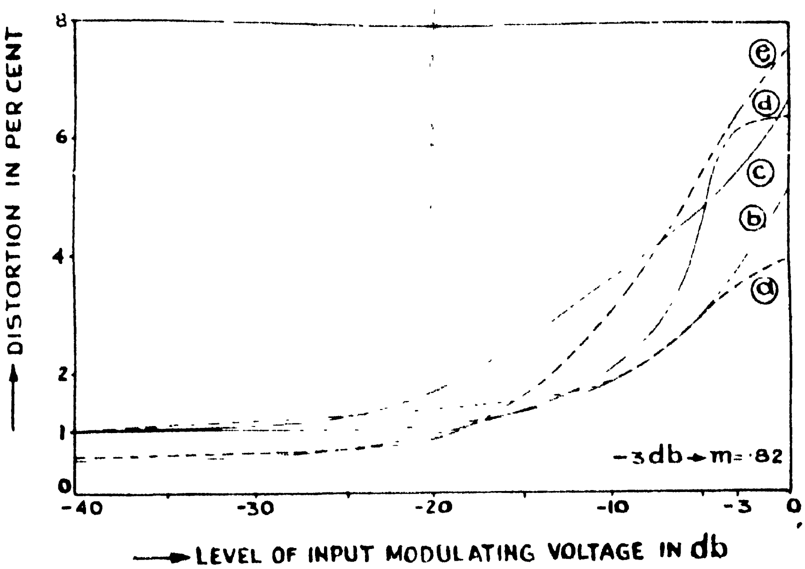

Fig. 5. Experimental total distortion with sliced inputs using 'Box-car' circuit for lowpass filters having different τ_2 and f_c.

$d = 10\mu S$, $tr_0 = 1.8\mu S$, Input level $= -3db$ for $m = 0.82$.

(a) With R-C l.p. filter ($6db$/oct); $\tau_2 = 0.5\mu sec$; $h_s = 0.5h_0$.

(b) With ,, ,, ,, ,, $\tau_2 = 1\mu sec$; $h_s = 0.36h_0$.

(c) With ,, ,, ,, ,, $\tau_2 = 3\mu sec$; $h_s = 0.15h_0$.

(d) With electronic l.p. filter, $f_c = 200$ Kc/s; $18db$/oct slope; $h_s = 0.6h_0$.

(e) With electronic l.p. filter; $f_c = 200$ Kc/s; $36db$/oct slope; $h_s = 0.6h_0$

III. CROSSTALK

Due to limited bandwidth, amplitude and phase distortion occur to the transmitted pulses and there is normally a carry-over of energy from one pulse to the following pulses. The crosstalk thus developed, may be caused either by high frequency cut-off or by low frequency cut-off. For high frequency cut-off with 6 db/oct. slope, simulated by an $R-C$ lowpass filter, the peak-to-peak carry-over voltage is found to be

$$C'_0(t) = h_0 \exp\left(-\frac{t}{\tau_2}\right)\left[(1+m)\left\{\exp\left(\frac{t_{r_{min}}}{\tau_2}\right)-1\right\} -(1-m)\left\{\exp\left(\frac{t_{r_{max}}}{\tau_2}\right)-1\right\}\right] \quad \ldots \quad (5)$$

where h_0 = mean height of the differentiated pulses.

τ_2 = RC of the lowpass filter.

The crosstalk ratio is then given by

$$\text{C.T. ratio} = \frac{2m \cdot \exp(t/\tau_2) \cdot \left[1-\exp\left(\frac{t_{r_{min}}}{\tau_2}\right)\right]}{\left[(1+m)\left\{\exp\left(\frac{t_{r_{min}}}{\tau_2}\right)-1\right\}-(1-m)\left\{\exp\left(\frac{t_{r_{max}}}{\tau_2}\right)-1\right\}\right]} \quad (6)$$

Since for high frequency cut-off, $C'_0(t)$ decreases rapidly with time, it is only necessary to consider the carry-over voltage from the channel pulse previous to the signal pulse. The numerical evaluation of Eq.(5) shows that $C'_0(t)$ is of opposite phase to that of the signal voltage. The experimental values of crosstalk ratios, with $d = 10\,\mu$ sec, $t_{r_0} = 1.82\,\mu$ sec and $m = 0.82$, are shown in Fig. 6 for various values of τ_2 and the channel separation Δt. These results agree closely with values obtained from Eq.(6).

For low frequency cut-off with 6 db/oct. slope, the crosstalk ratio for the carry-over voltage from the single previous channel (τ_3 small) is given by :

$$\text{C.T.ratio} = \frac{2m \cdot \exp^{\,t-t_{r_{min}}}}{\left[(1+m)\left\{1-\exp\left(\frac{t_{r_{min}}}{\tau_3}\right)\right\}-(1-m)\left\{1-\exp\left(\frac{t_{r_{max}}}{\tau_3}\right)\right\}\right]}$$

where τ_3 = RC of the highpass filter.

The carry-over vltage in this case is found to be of the same phase as that of the signal voltage. The experimental results with small values of τ_3 are also shown in Fig. 6 and they agree sufficiently with the results of Eq. (7).

As the coupling and decoupling circuits of pulse amplifiers are generally made large, the carry-over voltages from other previous channels have also to be considered to determine the net crosstalk ratio. With certain simplifying assumptions, it is found that the crosstalk ratio now is given by

$$\text{C.T. ratio} = \frac{2m\,\omega_m \tau_3 T_0}{[(1-m)t_{r_{max}}-(1+m)t_{r_{min}}]} \quad (8)$$

where ω_m = modulating angular frequency $\lll 1/T_0$; and $T_0 \lll \tau_3$. This result is similar to that obtained in case of P.A.M. (Flood, 1951). The carry-over voltage now is of opposite phase to that of the signal voltage.

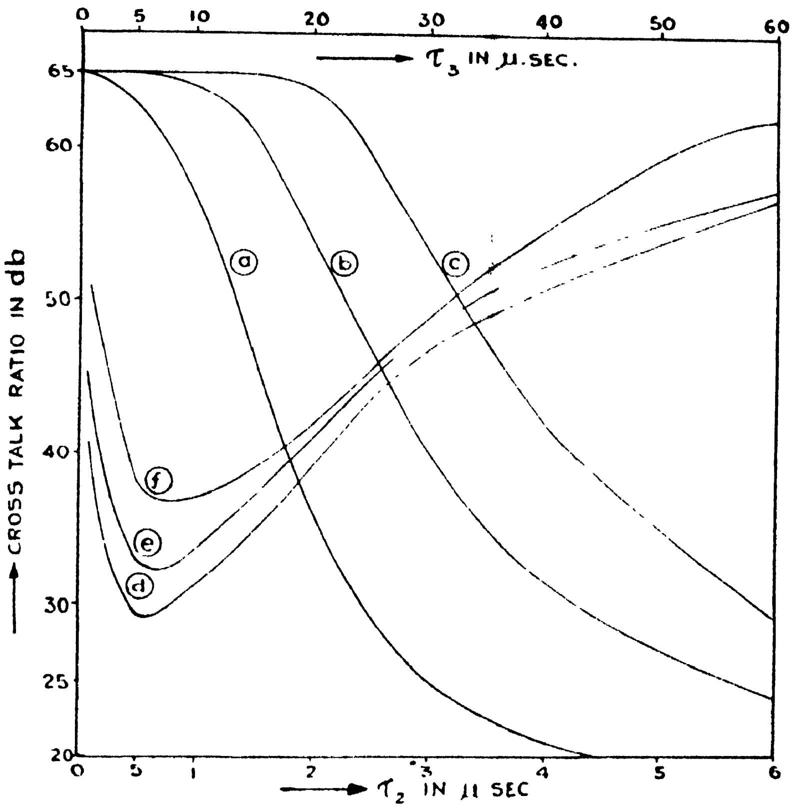

Fig. 6. Experimental crosstalk ratios with R–C filters for various τ_2, τ_3 and channel separation Δt.

d = 10 μS, tr_0 = 1.8 μS, m = 0.82, Input level = $-3db$.

(a) For R–C l.p. filter (τ_2) (6db/oct), Δt = 5 μ sec.

(b) " " " " , Δt = 10 μ sec.

(c) " " " " , Δt = 15 μ sec.

(d) For R–C h.p. filter (τ_3) (6db/oct), Δt = 5 μ sec.

(e) " " " " , Δt = 10 μ sec.

(f) " " " " , Δt = 15 μ sec.

Since with reference to the phase of the signal voltage, the carry-over voltage due to h.f. cut-off is of opposite phase and that due to l.f. cut-off with small values of τ_3 is of same phase, it is possible to minimise crosstalk by using simultaneous h.f. and l.f. cut-offs. The crosstalk ratio is now found to be

C.T. ratio

$$= \frac{2m\left[\exp\left(-\dfrac{t_{r_{min}}}{\tau_3}\right) - \exp\left(-\dfrac{t_{r_{min}}}{\tau_2}\right)\right]}{\left\{(1+m)\left[\exp\left(-\dfrac{t}{\tau_3}\right)\left\{1-\exp\left(\dfrac{t_{r_{min}}}{\tau_3}\right)\right\} - \exp\left(-\dfrac{t}{\tau_2}\right)\left\{1-\exp\left(\dfrac{t_{r_{min}}}{\tau_2}\right)\right\}\right] - (1-m)\left[\exp\left(-\dfrac{t}{\tau_3}\right)\left\{1-\exp\left(\dfrac{t_{r_{max}}}{\tau_3}\right)\right\} - \exp\left(-\dfrac{t}{\tau_2}\right)\left\{1-\exp\left(\dfrac{t_{r_{max}}}{\tau_2}\right)\right\}\right]\right\}} \quad \ldots (9)$$

The C.T. ratio will be maximum for certain relative values of τ_2 and τ_3. It is also possible to forecast the relative values of matching τ_2 and τ_3 for a certain channel separation Δt, as is given by

$$\Delta t = \left[\frac{\tau_2\tau_3}{\tau_3-\tau_2} \cdot \ln\left(\frac{D-C}{A-B}\right)\right] - [t_{r_{max}} + t_{r_{min}}] \qquad (10)$$

where,

$$A = \frac{(1+m)}{\tau_3}\left[\exp\left(\frac{t_{r_{min}}}{\tau_3}\right) - 1\right]$$

$$B = \frac{(1-m)}{\tau_3}\left[\exp\left(\frac{t_{r_{max}}}{\tau_3}\right) - 1\right]$$

$$C = \frac{(1+m)}{\tau_2}\left[1 - \exp\left(\frac{t_{r_{min}}}{\tau_2}\right)\right]$$

$$D = \frac{(1-m)}{\tau_2}\left[1 - \exp\left(\frac{t_{r_{max}}}{\tau_2}\right)\right]$$

Experimental results with simultaneous h.f. and l.f. cut-offs are shown in Fig. 7 and they agree closely with the results of Eq.(9) and of Eq. (10).

The pulse response with sharp cut-off filters is oscillatory (Guillemin, 1935) and the peak-to-peak carry-over voltage, for uniform transmission upto ω_c, is given by

$$C'_0(t) = \frac{h_0}{\pi}\{2m \,.\, Si(\omega_c t') - (1+m) \,.\, Si\left[\omega_c(t'-t_{r_{min}})\right] + (1-m) \,.\, Si\left[\omega_c(t'-t_{r_{max}})\right]\} \quad \ldots \quad (11)$$

where,

$$t' = t - t_d - (\Delta t + t_{r_{max}} + t_{r_{min}});$$

$-t_d\omega$ = phase shift within the pass band;

$$t_d \cong \pi/\omega_c$$

The crosstalk ratio is now $[4mh_0 f_c . t_{r_{min}}/C'_0(t)]$. The oscillatory nature of the C.T. ratio is shown in the experimental results of Fig. 8, obtained with an electronic filter.

In case of highpass sharp cut-off filters, the crosstalk ratio is given by

$$\text{C.T. ratio} = \frac{2m\pi}{\{(1+m) \,.\, Si[\omega_c(t'-t_{r_{min}})]-(1-m) \,.\, Si[\omega_c(t'-t_{r_{max}})]-2m \,.\, Si(\omega_c T')\}} \qquad \ldots \quad (12)$$

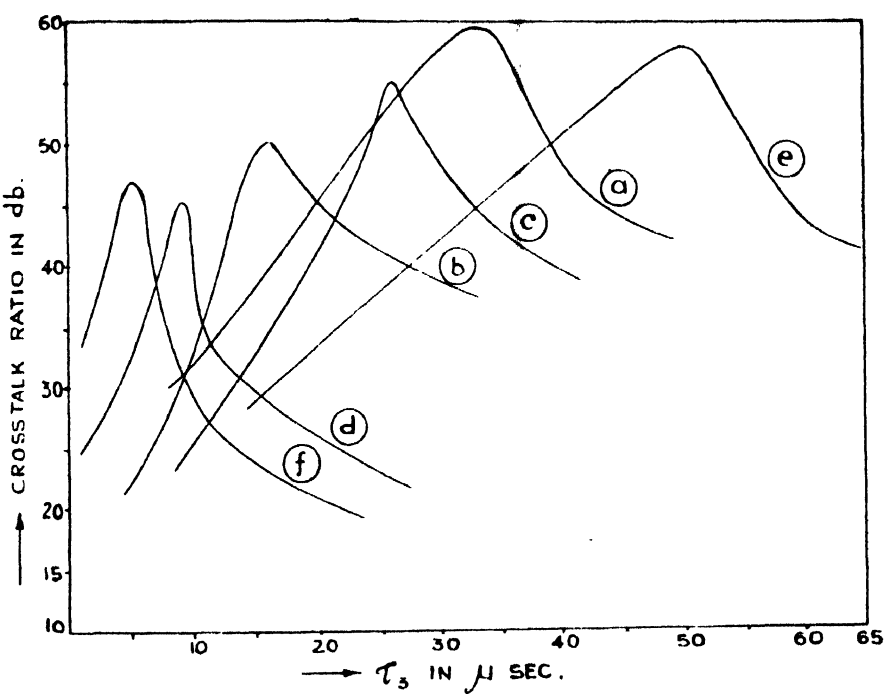

Fig. 7. Experimental crosstalk ratios due to combined H.F. and L.F. cut-off.
τ_2 = R–C of l.p. filter; τ_3 = R–C of h.p. filter; d = 10 μS, m = 0.82.

(a) $\Delta t = 5\ \mu S$; $\tau_2 = 1 \mu S$ (d) $\Delta t = 10\ \mu S$; $\tau_2 = 6\ \mu S$

(b) $\Delta t = 5\ \mu S$; $\tau_2 = 2\mu S$ (e) $\Delta t = 15\ \mu S$; $\tau_2 = 3\ \mu S$

(c) $\Delta t = 10\ \mu S$; $\tau_2 = 3\mu S$ (f) $\Delta t = 15\ \mu S$; $\tau_2 = 8\ \mu S$

Here the oscillations start only after the cut-off frequency becomes comparable with the pulse repetition frequency. Since the carry-over voltage due to both highpass and lowpass filters are oscillatory, it is possible to minimise crosstalk by simultaneous use of h.f. and l.f. cut-offs. In Fig. 8, certain minima on the C.T. ratio-curve could be improved to the points X_1', X_2', ... etc. by the use of highpass filters having suitable cut-off frequencies. The values of the cut-off frequencies f_{c_2} for the highpass filters are indicated on the same figure.

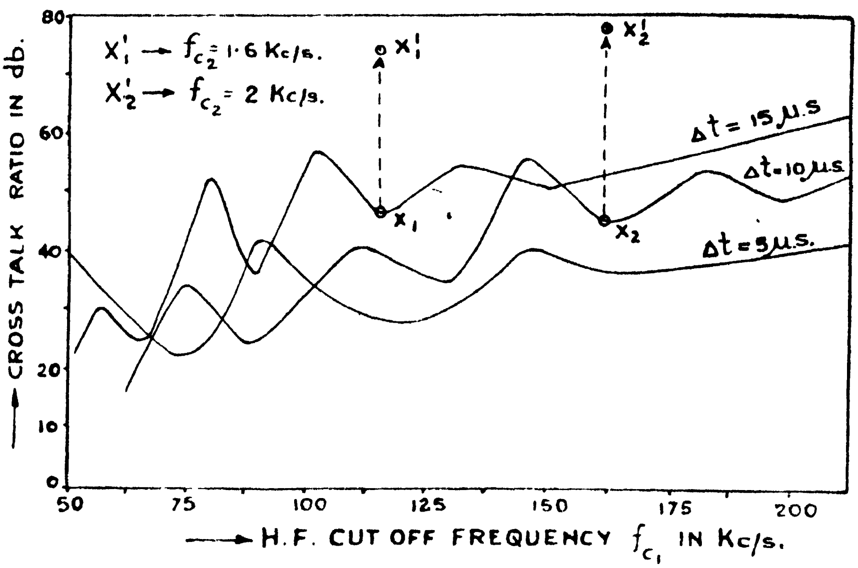

Fig. 8. Experimental crosstalk ratios due to electronic lowpass filter having $18db$/oct. slope.

$d = 10\mu S$; $m = 0.82$: f_{c_1} = cut-off frequency of l.p. filter.

f_{c_2} = cut-off frequency of h.p. filter.

X'_1 = Improved C.T. ratio with simultaneous use of l.p. and h.p. filters, $f_{c_2} = 1.6$ Kc/s.

X'_2 = Improved C.T. ratio with simultaneous use of l.p. and h.p. filters, $f_{c_2} = 2$ Kc/s.

When the slicer circuit is used, the slopes contain whatever carry-over voltages already present and the only difference is in the time of differentiation. Due to slicing, the differentiated pulses correspond to the slopes at a time later than $t = 0$, but the effective crosstalk ratios are not affected.

IV. NOISE

In case of random noise, the effective noise modulation with sharp cut-off media has been shown to be (Das, 1956)

$$(\Delta \tan \theta_N)_{eff.} = 0.1148\, f_c \times N(\text{peak}) \qquad \ldots \quad (13)$$

where, $\Delta \tan \theta_N = (\tan \theta_N - \tan \theta_{(mean)})$

$N_{(peak)}$ = peak amplitude of noise pulses

f_c = cut-off frequency.

The total audio noise power accepted by the a.f. amplifier is then proportional to

$$\frac{(\Delta \tan \theta_N)^2{}_{eff} \cdot F_a}{f_c} \times \left[\frac{tr_o^2}{T_0^2} + \left(\frac{2}{\pi^2}\right) \cdot \sum_{n=1}^{T_0 f_c} \frac{1}{n^2} \cdot \sin^2\left(\frac{\pi n tr_o}{T_0}\right) \right] \quad \dots \quad (14)$$

Numerical evaluation of Eq.(14), with $tr_o = 1\,\mu$ sec, $d = 10\,\mu$ sec, $m = 0.9$, $f_c = 1$ Mc/s and F_a = audio passband = 0 to 3.5 Kc/s, shows that

$$\frac{\text{R.M.S. audio signal}}{\text{Effective audio noise}} \text{ (in the output)} = \frac{1}{(1.414 T_0)} \cdot \frac{mA}{0.0956(\Delta \tan \theta_N)_{eff.}} \times \left(\frac{f_c}{F_a}\right)^{\frac{1}{2}} \quad \dots \quad (15)$$

$$= 10.9 \cdot \frac{A}{N_{(peak)}} .$$

This gives an output signal-to-noise ratio in db equal to $\left[20.74 + 20 \log \frac{A}{N_{(peak)}}\right]$.

For the extreme value of $\frac{A}{N_{(peak)}}$ equal to 2, the output signal-to-noise ratio is 26.74 db. After this threshold point, the improvement in the output S/N ratio in db is constant, but approximately varies as $(f_c/F_a)^{\frac{1}{2}}$. The theoretical and experimental results agree favourably as is shown in Fig. 9. The results of the impulse-noise tests show a further improvement in the output S/N ratios.

With transmission media having slower rate of cut-off, the equivalent bandwidth (Cherry, 1949) is determined and the above method is used for calculating the output S/N ratio. For simple $R-C$ lowpass filters, the equivalent cut-off frequency f_c is $1/4\tau$, and the minimum risetime is 2τ. To obtain a bandwidth of 1Mc/s, the time constant τ has to be 0.25 μ sec only.

V. DISCUSSION

Distortion in the P.S.M system has been very much minimized by using 'Box-car' pulse-lengthener circuit. The average distortion does not exceed 2% for media with 6 db/oct. cut-off rate. But for sharp cut-off media, the distortion is up-to 5% with larger modulation index. Distortion due to l.f. cut-off is less for all cases. Linearity of modulation and distortion characteristics are poorer when the slicer circuit is used. For higher level of modulation, the percentage distortion exceeds 5% at optimum slicing levels. But in case of P.L.M., Kretzmer (1947) has shown that even with large video bandwidths and ideal filters and amplifiers, the audio distortion is of the order of 4%. Levy (1949) has reported a P.P.M. system, where the overall distortion in audio characteristics was of the order of 5%. With bandwidth restriction, the distortion increases in P.L.M. and with $\tau_2 = 2\ \mu$ sec, the total distortion is found to be approximately 8% for similar pulse-width and repetition frequency.

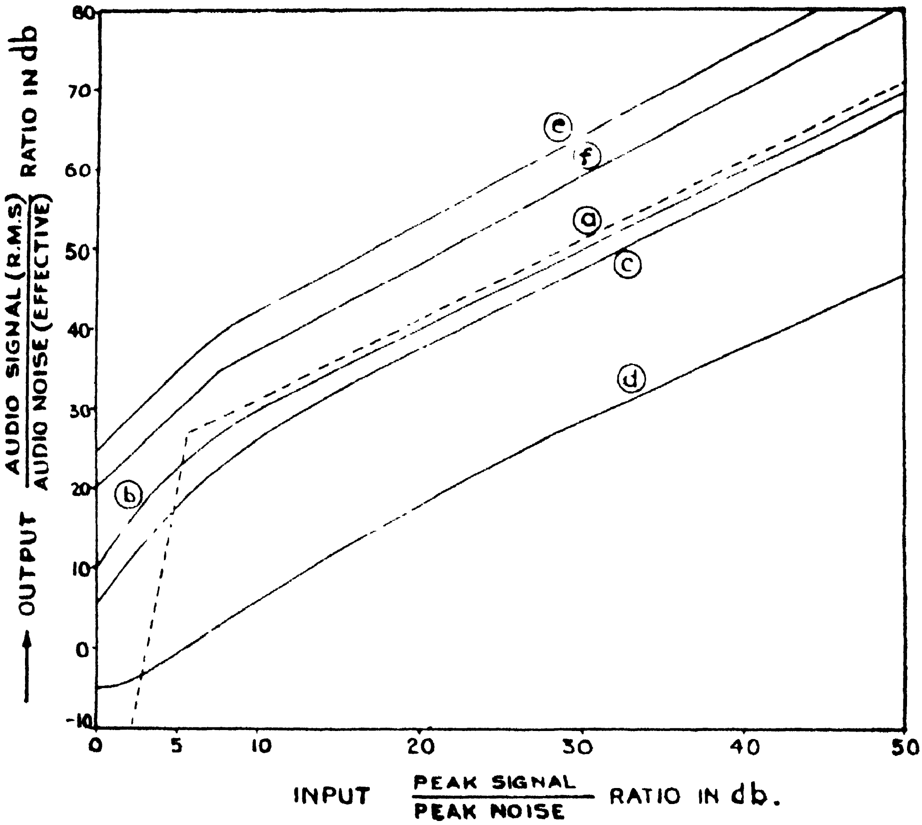

Fig. 9. Improvements in signal-to-noise ratio with sharp cut-off filters using slicers. $d = 10\ \mu$S, $F_a = 3.5$ Kc/s, P.R.F. = 10 Kc/s.

(a) Theoretical S/N ratio with random noise, $f_c = 1$ Mc/s, $t_{r_0} = 1\ \mu$ sec.

(b) Experimental ,, ,, ,, ,, ,, ,, ,,

(c) ,, ,, ,, ,, ,, $f_c = 0.5$ Mc/s; $t_{r_0} = 1.82\ \mu$S.

(d) ,, ,, ,, ,, ,, without slicer, $f_c = 1$ Mc/s.

(e) Experimental S/N ratio with impulse noise, $f_c = 1$ Mc/s; $t_{r_0} = 1\ \mu$S, impulse width = 1μS;, P.R.F. of impulses = 500 c/s.

(f) Experimental S/N ratio with impulse noise, $f_c = 1$ Mc/s; $t_{r_0} = 1\mu$S, impulse width = 1μS; P.R.F., of impulses = 5 Kc/s.

The other advantage of the P.S.M. system is in its better crosstalk characteristics. For a P.A.M. system with same pulse duration, the crosstalk ratio with $R-C$ lowpass filters is given as $\left(8.686 \times \frac{t-d}{\tau_2}\right) db$. In case of P.L.M. and P.P.M. (Deloraine, 1944) there is an approximate improvement factor of $\left(\frac{2t_m}{t_r}\right)$, where t_m is the time displacement due to modulation. From above, the improvement in P.S.M. crosstalk ratios are 10 to 17 *db* over those of P.A.M. and with a similar P.T.M. system, the crosstalk ratios would be at least equal.

In case of $R-C$ highpass filters, Flood (1951) has shown that the crosstalk ratio in P.A.M. is $\left(\frac{\omega_m T_0 \tau_3}{d}\right)$ for large τ_3 and small ω_m. Comparing with Eq. (8), the improvement in the P.S.M. crosstalk ratio with reference to P.A.M. is given by

$$\frac{\text{C.T. ratio in P.S.M.}}{\text{C.T. ratio in P.A.M.}} = \frac{2md}{|(1-m)t_{r_{max}} - (1+m)t_{r_{max}}|} = 44.3\ db. \qquad \ldots \quad (16)$$

for $\quad d = 2\ \mu\,\text{sec}; \quad t_{r_{max}} = 10\ \mu\,\text{sec}; \ t_{r_{max}} = 1\ \mu\text{sec}; \quad m = 0.82.$

Even with the improvements obtained in P.P.M. and P.L.M. (Flood, 1952) the P.S.M. crosstalk ratios would be better.

In Fig. 9, it is seen that above 10 *db* input S/N ratio, the experimental values are only 1.4 *db* below the theoretical values and there is an apparent threshold point at 7 *db* input S/N ratio for 1 Mc/s bandwidth. In a P.L.M. system with same maximum pulse duration and bandwidth, the improvement (Das, 1955) in the output S/N ratio is approximately 33 *db* over the input S/N ratio, whereas in P.S.M., the improvement is only about 21 *db*. However, the P.S.M. noise characteristics are definitely superior to those of P.A.M. and F.M. systems.

As an example of the overall performance of P.S.M., with filter-like transmission media, it is seen from Fig. 8, that the crosstalk ratio could be improved to about 80 *db* with $f_{c_1} = 160$ Kc/s and $f_{c_2} = 2$ Kc/s, and the corresponding distortion from Fig. 4 is only $1.5\% + 1\%$ for the highpass filter to be used in tendem. The improvement in the output S/N ratio is now only 16.5 *db*. If, then, the input S/N ratio is about 25 *db*, the total noise plus distortion in the output of the system will be about 3% only. This is a considerable advantage of the P.S.M. system.

Fourier expansion of the trapezoidal pulses shows that the useful modulated harmonic amplitudes vary as $1/n^2$, whereas for rectangular pulses the harmonic amplitudes vary as $1/n$, n being the order of the harmonic of the p.r.f. The P.S.M. signal which consists of trapezoidal pulses will then require a lesser nominal bandwidth for transmission. As the noise characteristics of P.S.M. are slightly inferior but the crosstalk and distortion characteristics are generally superior to those of P.T.M. systems for smaller bandwidths, P.S.M. will be more useful in low-noise bandwidth-limited systems like partially compensated cables and lines, electronic exchanges and others. If the noise level is low, the slicing level will be low and there will be very little audio distortion in the output.

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